

Article

Operation Modes of a Secondary-Side Phase-Shifted Resonant Converter

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Abstract: In this work, a unidirectional full-bridge resonant converter with secondary-side phase-shifted control is addressed. With the variation of voltage gain and power level, this converter may enter into different operational modes, including three continuous current modes and one discontinuous current mode. The details in each operation mode are discussed one by one. The boundary conditions between these modes have been identified. Additionally, the steady-state analysis of all continuous current modes is performed uniformly by using the fundamental harmonics approximation approach. Experimental results based on a lab prototype converter are also included to verify the operation modes predicted by the theoretical analysis.

Keywords: resonant converter; soft-switching; DC-DC converter

1. Introduction

Phase-shift modulation is a widely-used control method in power converters, especially in the bridge-type converters. The inner-bridge phase-shift refers to the adjustable phase-shift between different bridge arms in a single-phase or multiple-phase bridge inverter. As a consequence, the resultant high-frequency (HF) AC voltage may have an adjustable pulse width [1–6], which is also referred to as pulse-width modulation (PWM). The inter-bridge phase-shift refers to the adjustable phase-shift between two or more different bridges. The different inverter bridges may be all located on the input side, so that the phase-shift control manages both the power flow and power sharing among those bridges [7–9]. In other cases, two different inverter bridges may be located at the input side and the output side acting as an inverter and a rectifier, respectively. For example, the inter-bridge phase-shift control used in a dual-active bridge (DAB) converter, which is often called single-phase-shift (SPS) control, is simple and easy to implement [10–22]. Also referred as secondary phase-shift control, the inter-bridge phase-shift was applied for a unidirectional dual-bridge DC/DC converter [23–27]. The converter, including a semi-active rectifier with secondary phase-shift control, was proposed in [23] firstly and then investigated further experimentally in [24], which is proven to be suitable for high output voltage applications. By using unidirectional switches on the semi-active rectifier, two secondary phase-shift-controlled converters were reported to be able to maintain soft switching for a wide range of input/output variation [25,26]. All converters above make use of a power inductor as the main power transfer device, which actually can be replaced with a resonant tank. Due to the resonance phenomenon, the resonant version of the converter, including a semi-active rectifier, may have some extra different features. In this work, the resonant version of the converter discussed in [23] is selected to be investigated. It will be shown that it has more operation modes than that of the non-resonant converter. The detail of each mode and the transition conditions between different modes will be presented, as well.

The rest of the paper is organized as follows. The working principle and different operation modes of the secondary-side phase-shifted resonant converter are discussed at first in Section 2. In Section 3, the steady-state analysis in all continuous current modes is performed using fundamental harmonics approximation (FHA) uniformly. Boundary points between different operation modes can be identified. Necessary conditions of soft switching operation for all switches are derived, as well. In Section 4, the experimental results on a 300-W prototype converter will be reported to validate the different operation modes predicted by the theoretical analysis.

2. Operation Modes of the Secondary-Side Phase-Shifted Resonant Converter

2.1. Principle of Operation

Figure 1 presents the circuit diagram of the secondary-side phase-shifted resonant converter with a series resonant LC (an inductor and a capacitor) tank. It contains two full bridges on the two sides of an HF transformer. Four fully-controlled switches are used in the input bridge, while the output bridge is constituted of two diodes and two fully-controlled switches. In each bridge arm, the two active switches are turned on and off complementarily with nearly a 50% duty cycle. Although there are two phase-shifts available to be used to regulate the output power, only the inter-bridge phase-shift ϕ is utilized for power control in this work, which is defined as the phase delay between gating signals of M_1 and M_6 . The inner-bridge phase-shift between the two arms of the input bridge is fixed at 180° . Hence, the output HF AC voltage of input bridge v_p is always a square wave voltage. On the other side, the input HF AC voltage of the output bridge (actually, the transformer secondary voltage) v_T is an irregular quasi-square waveform whose shape depends on ϕ , the power level and converter specifications. This is attributed to the resonant feature and the existence of the diode arm in the output bridge.

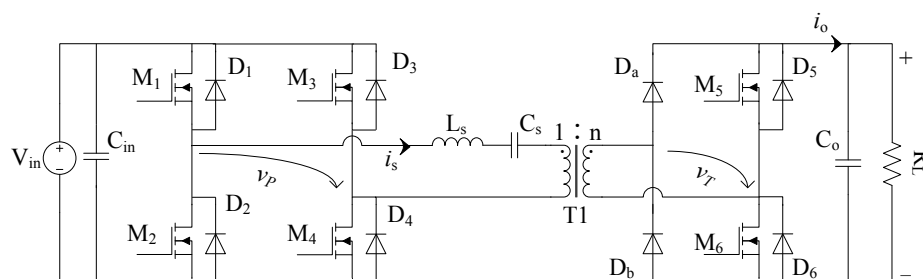


Figure 1. A secondary-side phase-shifted resonant converter.

Based on the continuity of the resonant current i_s in the steady state, two possible kinds of operation modes may exist: continuous current mode (CCM) and discontinuous current mode (DCM). Through careful investigation, it is found that there are in total three different CCMs and one DCM in which the converter may work. In each mode, the involved time intervals in each HF switching period are different. In the following part, the three CCMs and one DCM will be discussed one by one. Two parameters are defined first to ease the explanation. The converter voltage gain is defined as: $d = V_o/nV_{in}$, where $1:n$ is the primary-to-secondary turns ratio of the HF transformer. The angle β is defined as the phase angle by which v_p leads the resonant current i_s . Effects of the deadband and snubber are neglected.

2.2. Continuous Current Mode 1 (CCM1)

As long as the phase-shift ϕ is large enough, the resonant converter can work in CCM1 regardless of the converter gain d . The signature of CCM1 is $\phi > \beta$. The typical waveforms in CCM1 is shown in Figure 2. In an HF period of CCM1, there are six different time intervals, whose equivalent circuits have been included in Figure 3.

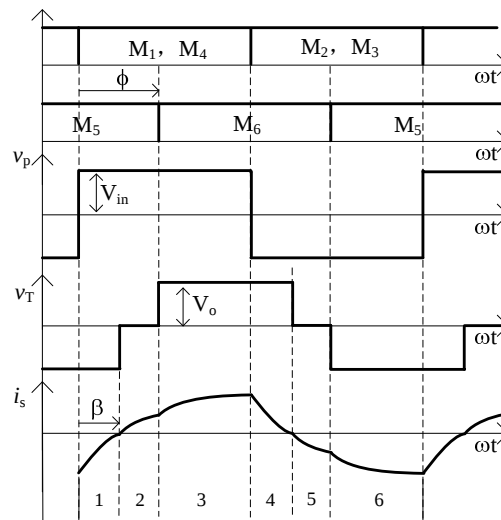


Figure 2. Steady-state waveforms of continuous current mode 1 (CCM1) operation with $\phi > \beta$.

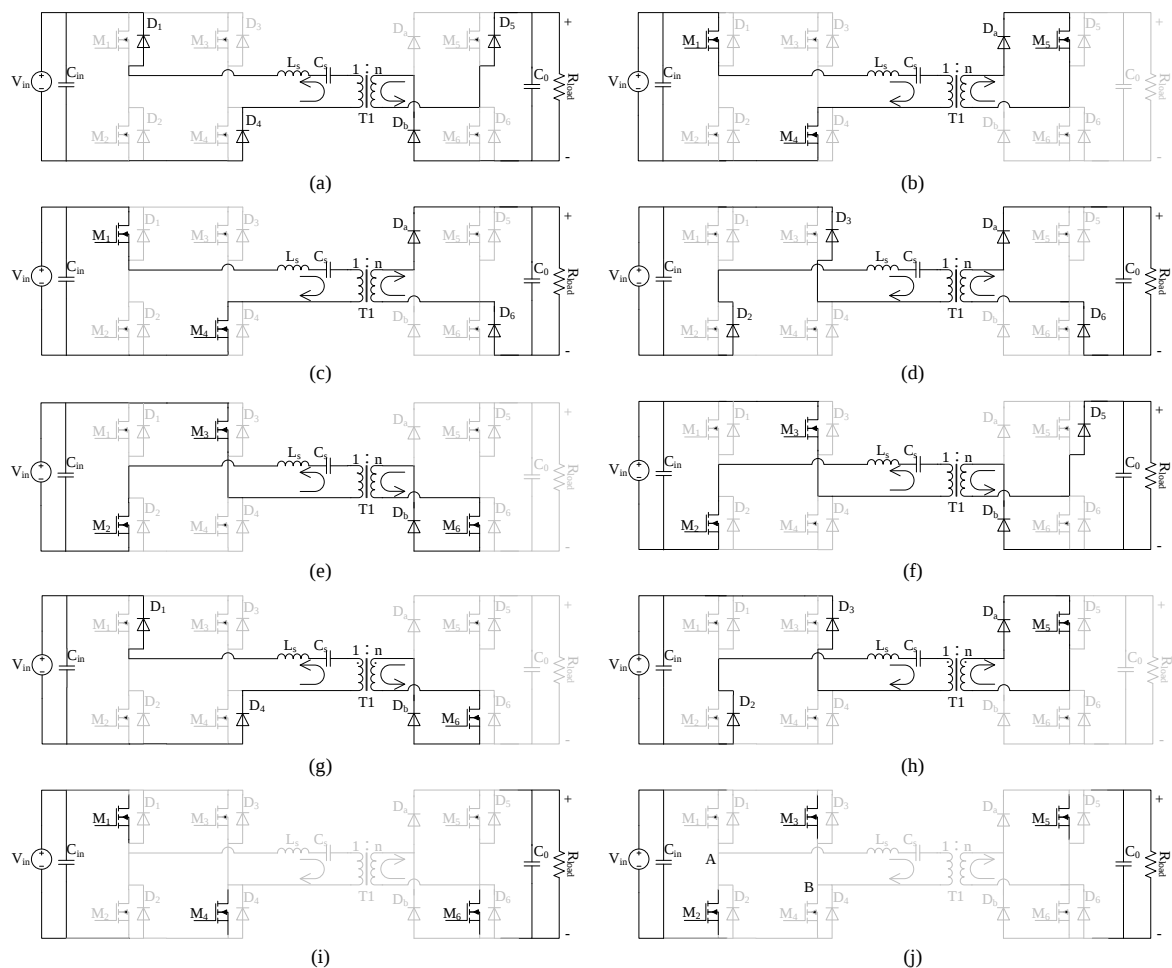


Figure 3. Equivalent circuits during different time intervals in a high-frequency (HF) period.

It is assumed that before the beginning of a new period, the resonant current i_s is negative, and switches M_2, M_3, M_5 are in the on state. Interval 1 (Figure 3a): This begins when the negative-to-positive transition of v_p is triggered by the turn-off of M_2 and M_3 . i_s is shifted from

M_2 , M_3 to D_1 , D_4 naturally. The conducting devices in this interval are D_1 , D_4 , D_b , D_5 . It ends when the resonant current reaches zero.

Interval 2 (Figure 3b): When i_s changes direction, the primary-side switches M_1 and M_4 are turned on with zero voltage. On the secondary side, D_b is turned off and D_a is turned on, both with zero current. M_5 is turned on with zero voltage, as well. Thus, the secondary side of the transformer is shorted by means of M_5 and D_a . As a result, the input voltage is solely applied on the resonant tank. The energy is being stored in the resonant tank. The interval ends when M_5 is turned off.

Interval 3 (Figure 3c): With the turn-off of M_5 , the secondary current is forced to flow in D_6 , and the gating signal of M_6 can be applied during this interval with almost zero voltage. The energy is pumped into the output through the resonant tank. This interval completes at the removal of the gating signals of M_1 , M_4 .

Intervals 4–6: The intervals during the next half of the HF switching cycle are almost the same as those described above, except for the conducting devices and the direction of the resonant current. The details will not be difficult to understand by referring to Figure 3d–f.

2.3. Continuous Current Mode 2 (CCM2)

For a converter with $d < 1$ working in CCM1, if the phase-shift ϕ is reduced continuously, the converter would enter into CCM2 operation eventually. The signature of CCM2 is $\phi < \beta$. Based on the variation of ϕ , there are three sub-modes in CCM2 for $\phi > 0$, $\phi = 0$, $\phi < 0$, which are shown in Figure 4a–c, respectively. The first case of $\phi > 0$ is taken as an example. Similar to CCM1, there are also six different time intervals in one HF period for this case.

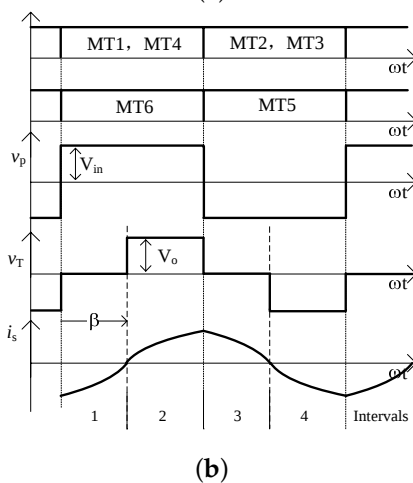
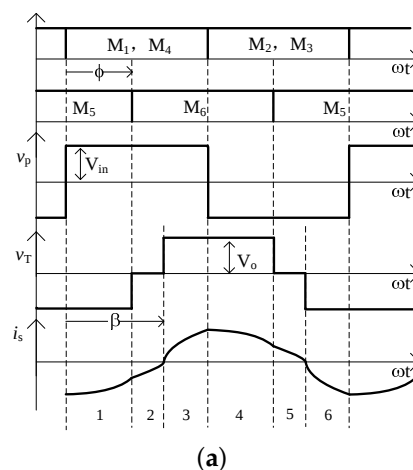


Figure 4. Cont.

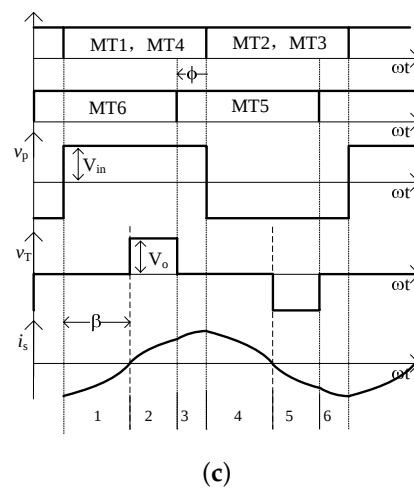


Figure 4. Steady-state waveforms of CCM2 operation with $\phi < \beta$: (a) $\phi > 0$; (b) $\phi = 0$; (c) $\phi < 0$.

Interval 1 (Figure 3a): This is almost the same as Interval 1 of CCM1, except that it ends when M_5 is turned off and M_6 is turned on with a negative secondary current.

Interval 2 (Figure 3g): With the turn-off of M_5 , the negative secondary current is forced to flow in the drain-to-source channel of M_6 , which is a lossy turn-on process. The operation conditions on the primary side have no change. It is also an energy-storage interval and will last until i_s declines to zero.

Interval 3 (Figure 3c): This is almost the same as the Interval 3 in CCM1, except that it starts with a zero resonant current. In other words, the primary resonant current is shifted from D_1, D_4 to M_1, M_4 naturally in the zero-voltage condition. The secondary current is shifted from D_b, M_6 to D_a, D_6 .

Again, the operations in Intervals 4–6 are almost the same as those described above, except for the conducting devices and the direction of the resonant current. Those can be understood by referring to Figure 3d,h,f.

The second case (Figure 4b) of CCM2 comes into being when $\phi = 0$, which only has four intervals in one HF period. The corresponding equivalent circuits are (g), (c), (h) and (f) in Figure 3. The third case (Figure 4c) of CCM2 comes into being when $\phi < 0$, which has six intervals in one HF period. The corresponded equivalent circuits are Figure 3g, c, b, h, f, e.

2.4. Just Continuous Current Mode (JCCM)

For a converter with $d > 1$ working in CCM1, if the phase-shift ϕ is reduced continuously, the converter would enter into one special operation mode: JCCM (just continuous current mode). The signature of JCCM is $\beta = 0$, i.e., the resonant current i_s is synchronous with v_p . JCCM is a special mode, which can be categorized as either CCM or DCM. There are only four time intervals in the JCCM operation, which are shown in Figure 5.

Interval 1 (Figure 3b): This is similar to Interval 2 of CCM1. The interval starts with the zero-current turn-on of M_1, M_4 and ends with the turn-off of M_5 . This is an energy-storage interval.

Interval 2 (Figure 3c): This is similar to Interval 3 in CCM1, except that it ends with the zero-current turn-off of M_1, M_4 . It is a powering interval.

Intervals 3~4: These two are symmetric operations of the first half period, which can be understood easily by referring to Figure 3e,f.

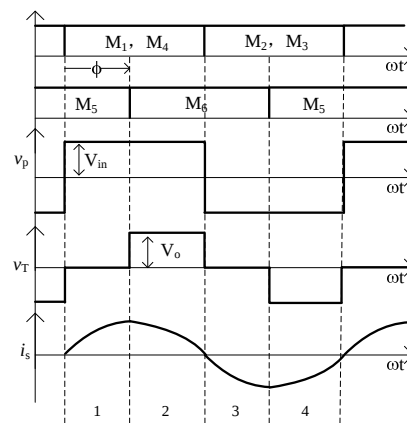


Figure 5. Steady-state waveforms of the just continuous current mode (JCCM) operation with $\beta = 0$.

2.5. Continuous Current Mode 3 (CCM3)

When the converter works in JCCM, reducing the phase-shift ϕ further may have two different consequences. One is CCM3 (Figure 6), and the other is DCM (Figure 7).

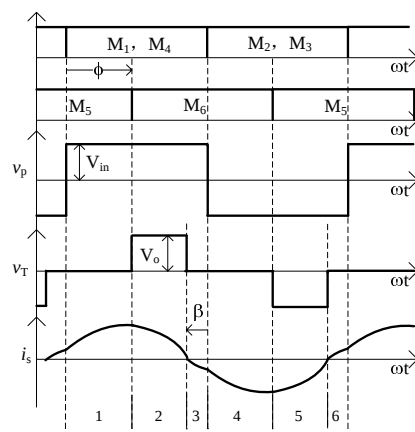


Figure 6. Steady-state waveforms of the CCM3 operation.

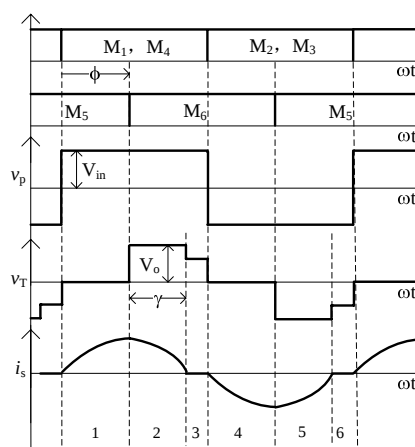


Figure 7. Steady-state waveforms of the discontinuous current mode (DCM) operation.

When the positive resonant current declines to zero, the current tends to reverse direction. Because there are no changes made to any gating signals, the direction reversal of resonant current will result in $v_T = 0$ if the current goes negative. According to Kirchhoff's voltage law, the instant resonant capacitor voltage must be larger than the instant value of v_P (which is V_{in} now) to let the current change from zero to negative. This consequence is the CCM3 operation. The CCM3 is unique to the resonant-type converter, which cannot be observed in the converter proposed by [23]. It is a conditional state and can only exist when the following conditions are met:

$$d > 1; v_{cp} > V_{in} \quad (1)$$

where v_{cp} is the peak voltage of the resonant capacitor.

The typical steady-state waveforms in CCM3 are shown in Figure 6. Interval 1 (Figure 3b): This is similar to Interval 2 of CCM1, except that M_1, M_4 are turned on with positive resonant current, which indicates the loss of ZVS.

Interval 2 (Figure 3c): This is similar to Interval 3 in CCM1, except that it ends when the resonant current declines to zero. It is a powering interval.

Interval 3 (Figure 3g): This is similar to Interval 2 in CCM2, except that it starts with the resonant current shifting from D_a, D_6 to D_b, M_6 . Additionally, it ends with the turn-off of M_1, M_4 .

Intervals 4~6: These three are symmetric operations of the first half period, which can be understood easily by referring to Figure 3e,f,h.

2.6. DCM

Only existing with $d > 1$, DCM is featured with a portion of zero-current duration in each half HF period. There are two possibilities for the converter to move into DCM. The first possibility is evolving from JCCM directly. A slight reduction of phase-shift ϕ from JCCM with $v_{cp} < V_{in}$ will result in DCM operation immediately. The other possibility is evolving from CCM3. With the decreasing of ϕ , the resonant current at the beginning of Interval 1 of CCM3 will decrease too, until it reaches zero. Thus, Interval 3 of CCM3 becomes an idle interval of DCM. When the positive resonant current reaches zero in DCM, it cannot go negative instantly, since $v_{cp} < V_{in}$. Thus, the resonant current will remain zero, and the resonant capacitor voltage is clamped at its peak value, as well. The transformer primary-side voltage is the difference of V_{in} and V_{cp} according to Kirchhoff's voltage law.

The typical steady-state waveforms in DCM is shown in Figure 7. The DCM operation is almost the same as JCCM with one idle interval inserted in each half period. In the idle interval (Figure 3i or Figure 3j), there is no current flowing in the tank and transformer. However, M_1, M_4, M_6 in Figure 3i are still on and are able to conduct current. Both D_a and D_b are reversely biased, and the reverse voltages are $V_{Da} = V_o - n(V_{in} - V_{cp})$; $V_{Db} = n(V_{in} - V_{cp})$.

3. Steady-State Analysis of Continuous Current Modes

In this section, the steady-state analysis for CCM operation of the resonant converter would be performed using the fundamental harmonics approximation (FHA) approach. The first reason to use FHA is that all AC signals in the converter are fundamental dominant signals. The second reason is that the analysis of all CCM modes could have a universal solution by using the FHA approach. For the DCM operation, the special design-dependent waveform of the transformer secondary voltage makes it hard to apply FHA. DCM analysis can be done in the time domain by solving differential equations precisely, which will be addressed in a separate work. It is assumed that all switches, including diodes and passive components, are ideal. Snubber effects and the switch commutation transient are neglected, as well. To facilitate the analysis, all parameters will be normalized by the following base values:

$$V_B = V_{in}; Z_B = R'_L; I_B = \frac{V_B}{Z_B}; \omega_B = \omega_r = \frac{1}{\sqrt{L_s C_s}} \quad (2)$$

where R'_L is the primary reflected full load resistance and ω_r is the resonance frequency.

The normalized values of all reactances are given by:

$$X_{Ls,pu} = QF; X_{Cs,pu} = -\frac{Q}{F} \quad (3)$$

$$X_{s,pu} = X_{Ls,pu} + X_{Cs,pu} = Q(F - \frac{1}{F}) \quad (4)$$

where:

$$F = \frac{\omega_s}{\omega_r}; \quad Q = \frac{\omega_r L_s}{R'_L} \quad (5)$$

are the normalized switching frequency and the quality factor, respectively; and ω_s is the switching frequency.

The fundamental equivalent circuit of the converter in the phasor domain is given in Figure 8. The two voltage sources are the normalized fundamental phasor of $v_p(t)$ and $v_T(t)/n$, respectively. Although there are many different CCM cases, the phase-shift between two fundamental voltage sources is found to have a unified form: $-\frac{\beta+\phi}{2}$ for all CCM cases. Thus, the two voltage phasors can be written as:

$$\mathbf{V}_{p,pu} = \frac{4}{\pi} \angle 0 \quad (6)$$

$$\frac{\mathbf{V}_{T,pu}}{n} = \frac{4d}{\pi} \cos \frac{\beta-\phi}{2} \angle -\frac{\beta+\phi}{2} \quad (7)$$

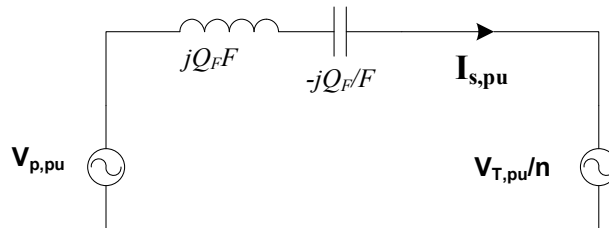


Figure 8. The fundamental equivalent circuit of the CCM operation in the phasor domain.

Then, the resonant current phasor can be easily obtained as:

$$\mathbf{I}_{s,pu} = \frac{\frac{4}{\pi} \angle 0 - \frac{4d}{\pi} \cos \frac{\beta-\phi}{2} \angle -\frac{\beta+\phi}{2}}{jQ(F - \frac{1}{F})} \quad (8)$$

After solving the equation, the normalized resonant current i_s in the time domain is shown as:

$$i_{s,pu}(t) = I_{sp,pu} \sin(\omega_s t - \beta) \quad (9)$$

where the peak resonant current $I_{sp,pu}$ and the phase angle β are given as:

$$I_{sp,pu} = \frac{4F \sqrt{1 + d^2 \cos^2 \frac{\beta-\phi}{2} - d(\cos \beta + \cos \phi)}}{\pi (QF^2 - Q)} \quad (10)$$

$$\beta = \arctan \frac{2 - d(\cos \beta + \cos \phi)}{d(\sin \beta + \sin \phi)} \quad (11)$$

By simplifying Equation (11), an important equation involving β , ϕ and d can be obtained as:

$$d = \frac{2 \cos \beta}{1 + \cos(\beta - \phi)} \quad (12)$$

This equation can be used to find β numerically if d and ϕ are known. Substituting Equation (12) into Equation (10) yields:

$$I_{sp,pu} = \frac{4F \sqrt{1 - d \cos \phi}}{\pi (QF^2 - Q)} \quad (13)$$

This equation demos the relationship among the normalized peak resonant current and the converter gain d , the phase-shift ϕ . The normalized peak voltage across the resonant capacitor can be obtained as:

$$V_{cp,pu} = \frac{4 \sqrt{1 - d \cos \phi}}{\pi (F^2 - 1)} \quad (14)$$

The transferred power from the primary side to the secondary side could be calculated from either side. Here, the instantaneous power is evaluated on the primary side:

$$\begin{aligned} p_{pu}(\omega_s t) &= v_{p,pu}(\omega_s t) \times i_{s,pu}(\omega_s t) \\ &= \frac{4}{\pi} \sin \omega_s t \times I_{sp,pu} \sin(\omega_s t - \beta) \end{aligned} \quad (15)$$

The normalized output power is the average value of instantaneous power:

$$P_{o,pu} = \frac{1}{2\pi} \int_0^{2\pi} p_{pu}(\omega_s t) d(\omega_s t) = P_m d (\sin \beta + \sin \phi) \quad (16)$$

where $P_m = \frac{4F}{\pi^2(QF^2 - Q)}$.

Besides, the converter gain d in the CCM operation can be evaluated by:

$$d = \frac{4F (\sin \beta + \sin \phi)}{\pi^2 (QF^2 - Q)} Z_L \quad (17)$$

where $Z_L = R_{Load} / (n^2 Z_B)$ is the normalized load resistance in the range of $1 \sim \infty$. According to the equation above, at a lighter load (Z_L increases), the sum of two sinusoidal terms in the numerator has to be reduced to keep the converter gain constant.

3.1. $d = 1$

When the converter gain is unity, the converter can only operate in CCM1. By letting Equation (12) equal one, the relationship between two angles is given as:

$$\phi = \beta + \arccos(2 \cos \beta - 1) \quad (18)$$

With ϕ reduced to get lower power, β will decline, as well. At no load condition, both ϕ and β are zero. The variations of $P_{o,pu} / P_m$ and β with regards to ϕ are presented in Figure 9, respectively.

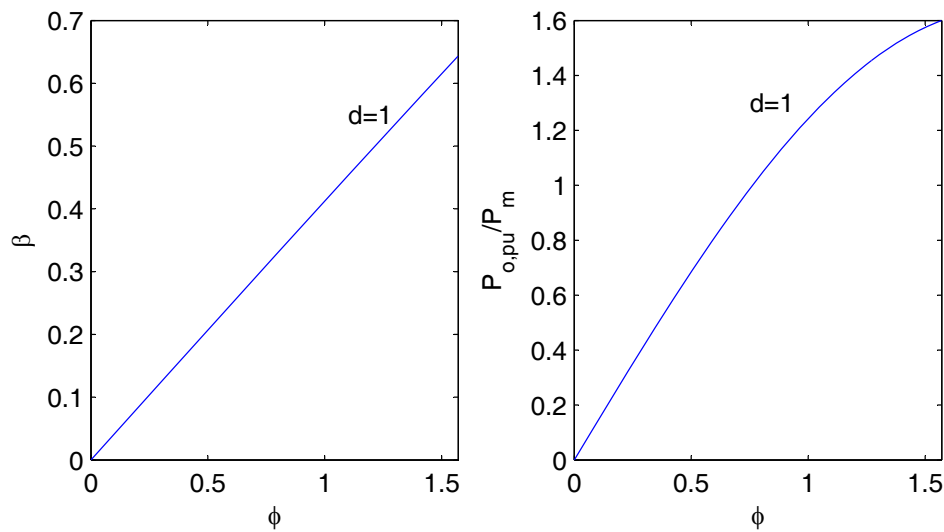


Figure 9. The variations of $P_{o,pu}/P_m$, β with regards to ϕ at $d = 1$.

3.2. $d < 1$

When the converter gain is less than one, the converter might operate in either CCM1 or CCM2, as shown in Figure 10. At high power with $d < 1$, the converter is working in CCM1 with $\phi > \beta > 0$. With the decreasing of the power level, β also goes down with a slower speed following ϕ . When those two angles equal each other, the converter is going to move into CCM2. By substituting $\beta = \phi$ into Equation (12), the boundary condition is obtained as:

$$d = \cos \phi = \cos \beta \quad (19)$$

The boundary point has been highlighted in Figure 10. The above equation also shows that CCM2 can happen only at $d < 1$, since a cosine term can never be larger than one. As indicated in Figure 10, β will increase if ϕ is further reduced from the value at the boundary point. ϕ may be reduced to zero and even negative in order to regulate power. In the no-load condition, ϕ approaches $-\pi/2$ and β approaches $\pi/2$.

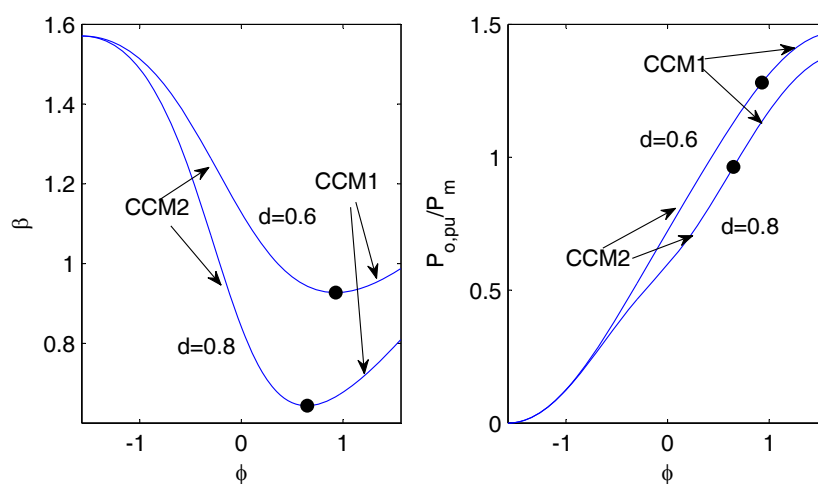


Figure 10. The variations of $P_{o,pu}/P_m$, β with regards to ϕ at $d < 1$.

3.3. $d > 1$

When the converter gain is larger than one, the converter might operate in one of CCM1, CCM3, JCCM and DCM. At high power with $d > 1$, the converter is working in CCM1 with $\phi > \beta > 0$. To achieve low output power, ϕ is reduced to keep converter gain. β will decrease too with a higher gradient. When $\phi > 0 = \beta$, the converter is working in JCCM. For JCCM, the relationship between ϕ and d is shown as:

$$d = \frac{2}{1 + \cos \phi_2}; \text{ or } \phi = \arccos\left(1 - \frac{2}{d}\right) \quad (20)$$

It can be seen clearly from Equation (20) that the JCCM operation can be attained only if $d > 1$.

As mentioned before, the resultant mode with an even smaller ϕ depends on the relationship between V_{cp} and V_{in} . By letting Equation (14) equal one when $\beta = 0$, it is seen that:

$$d^* = \frac{\pi^2(F^2 - 1)^2}{16} + 1 \quad (21)$$

If $d \leq d^*$, the converter can go into DCM immediately once ϕ is reduced from the value in JCCM; otherwise, the converter will enter into CCM3 first and then DCM. The boundary point between CCM3 and DCM is given as:

$$\phi = \arccos\left(\frac{2 - d^*}{d}\right) \quad (22)$$

In Figure 11, the variations of $P_{o,pu}/P_m$, β with regards to ϕ at $d > 1$ are presented. As an example, $F = 1.1$ is used to calculate d^* , which is a common selection in the literature. With the help of Equation (21), the critical gain is found to be $d^* = 1.027$. Thus, three cases for $d < d^*$, $d = d^*$, $d > d^*$, respectively, are illustrated in Figure 11. All of the boundary points between each mode are highlighted in the plots.

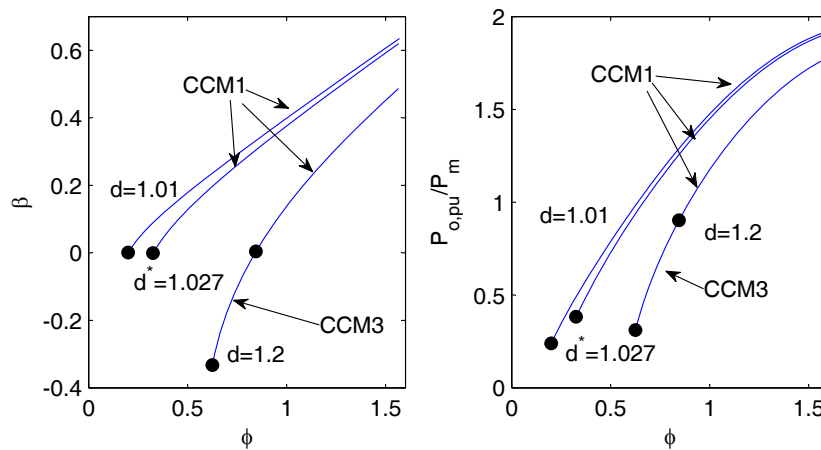


Figure 11. The variations of $P_{o,pu}/P_m$, β with regards to ϕ at $d > 1$.

3.4. Soft Switching Conditions

The soft switching conditions of all switches and diodes in the converter can be determined by the polarity of the current during the switching transition. The value of the transition current can be found by evaluating Equation (10) at different transition points. For example, a negative resonant current at $\omega_s t = 0$ enables switches $M_1 \sim M_4$ to work with ZVS possibly. Additionally, a positive resonant current at $\omega_s t = \phi$ is necessary for M_5 , M_6 to work with ZVS. In Table 1, the soft-switching behavior in the converter for different operation modes is listed. It can be concluded that only the CCM1 operation is preferable in terms of almost zero switching loss.

Table 1. Soft switching conditions for different operation modes.

Modes	ZVS	ZCS (zero-current switching)	Hard switching
CCM1	$M_1, M_2, M_3, M_4, M_5, M_6$	D_a, D_b	-
CCM2	M_1, M_2, M_3, M_4	D_a, D_b	M_5, M_6
CCM3	M_5, M_6	D_a, D_b	M_1, M_2, M_3, M_4
DCM	M_5, M_6	$M_1, M_2, M_3, M_4, D_a, D_b$	-

4. Validation by Experimental Test

4.1. Design Point

The analysis results in the last section can be used for a quick initial design of a secondary-side phase-shifted resonant converter. It is assumed that the converter to be investigated has the nominal specifications as: $V_{in} = 110$ V, $V_o = 100$ V, $f_s = 100$ kHz, $P_o = 300$ W. Usually the normalized switching frequency F is chosen to be larger than one, but close to one to achieve an inductive resonant current for the ZVS operation in the input bridge and will not bring up large circulation current at the same time. A large F will also result in a resonant tank with a large size. Hence, $F = 1.1$ is selected in this design based on experience [20]. The quality factor Q is proportional to the size of the resonant inductor and has an effect on the rms resonant current. $Q = 1$ is chosen for the purpose of illustration. Since the converter with $d = 1$ may work with ZVS for a wide range of load levels, the converter gain at nominal input voltage is selected at unity.

Therefore, the HF transformer turns ratio is calculated as: $1:n = dV_{in}:V_o = 11:10$. Additionally, the resonant tank can be found as:

$$L_s = \frac{QFd^2V_{in}^2}{2\pi f_s P_o} = 70.6 \mu\text{H} \quad (23)$$

$$C_s = \frac{FP_o}{2\pi f_s Qd^2V_{in}^2} = 43.4 \text{ nF} \quad (24)$$

4.2. Experimental Results

According to the initial design, a lab prototype converter is built and tested to validate the operation modes predicted theoretically. The specifications of the prototype are listed in Table 2. The converter is tested under different input voltages and different load levels. The test waveforms are shown in Figures 12–18.

Table 2. The specifications of the prototype.

Parameters	Value
Nominal input voltage	110 V
Maximum input voltage	120 V
Minimum input voltage	80 V
Output voltage	100 V
Resonant inductor	70.6 μH
Resonant capacitor	43.4 nF
HF transformer	11:10, core size: ETD39, material: N97
Switching frequency	100 kHz

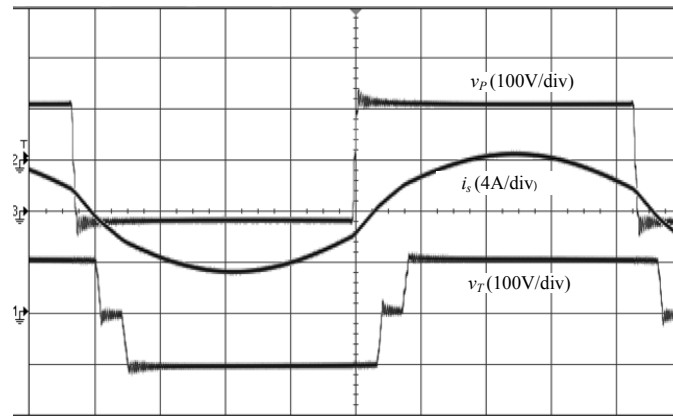


Figure 12. Experimental plot at: $d = 1$; $V_{in} = 110$ V; $V_o = 100$ V; $P_o = 300$ W, CCM1 operation.

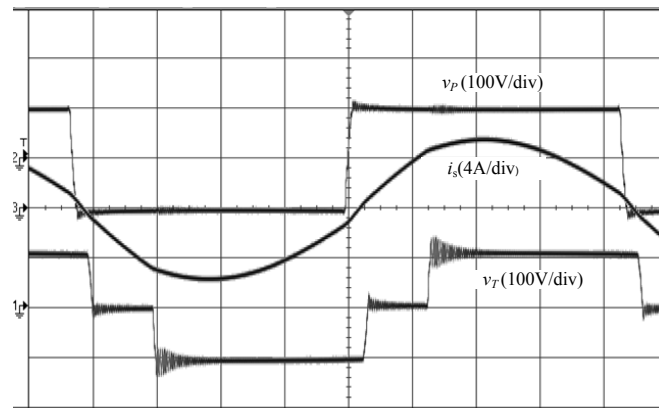


Figure 13. Experimental plot at: $d > 1$; $V_{in} = 100$ V; $V_o = 100$ V; $P_o = 300$ W, CCM1 operation.

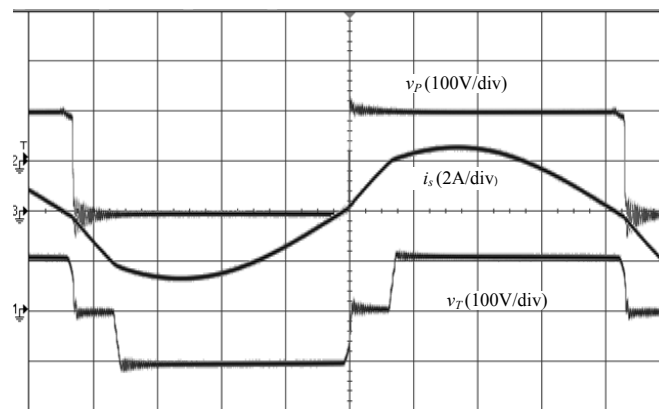


Figure 14. Experimental plot at: $d > 1$; $V_{in} = 100$ V; $V_o = 100$ V; $P_o = 150$ W, nearly JCCM operation.

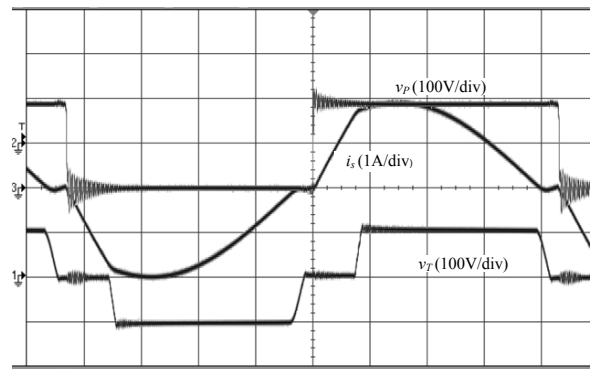


Figure 15. Experimental plot at: $d > 1$; $V_{in} = 90$ V; $V_o = 100$ V; $P_o = 90$ W, CCM3 operation.

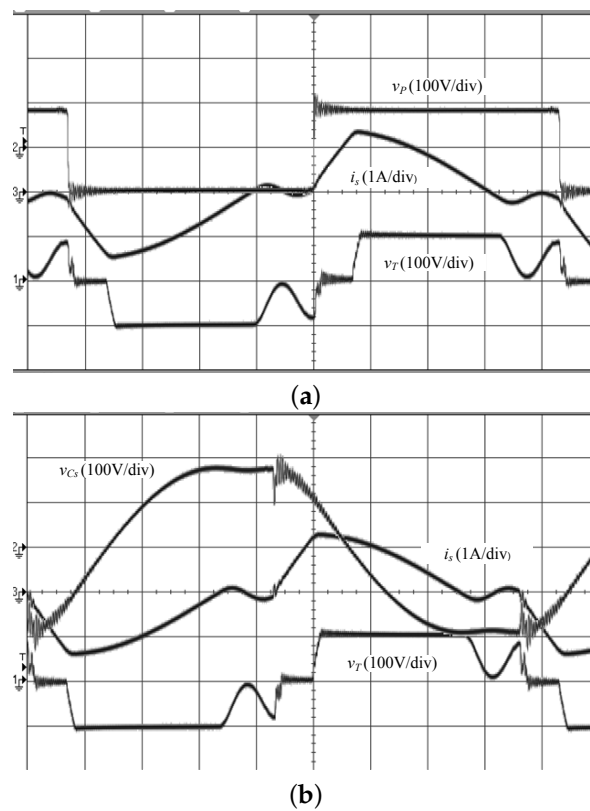


Figure 16. Experimental plot at: $d > 1$; $V_{in} = 80$ V; $V_o = 100$ V; $P_o = 45$ W, DCM operation. (a) the HF inverter output voltage v_p , the resonant current i_s and the transformer voltage v_T ; (b) the resonant capacitor voltage v_p , the resonant current i_s and the transformer voltage v_T .

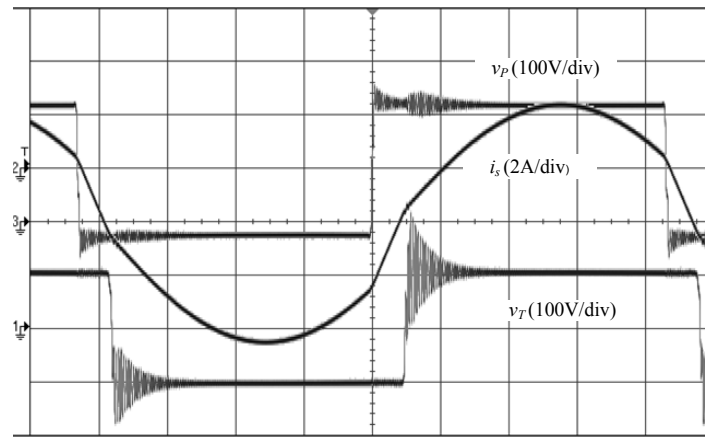


Figure 17. Experimental plot at: $d < 1$; $V_{in} = 120$ V; $V_o = 100$ V; $P_o = 270$ W, CCM2 operation with $\phi > 0$.

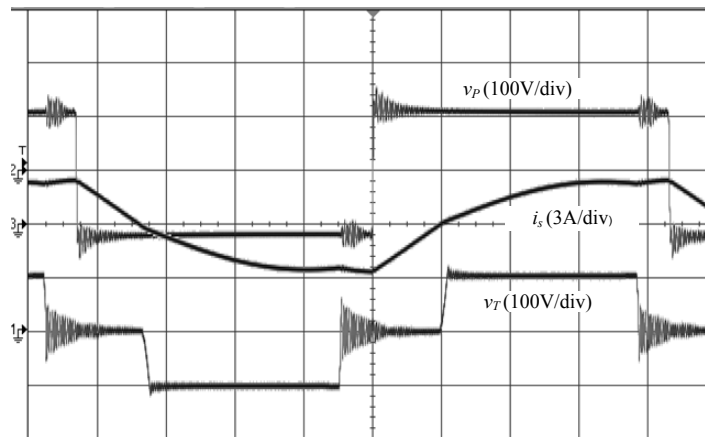


Figure 18. Experimental plot at: $d < 1$; $V_{in} = 120$ V; $V_o = 100$ V; $P_o = 120$ W, CCM2 operation with $\phi < 0$.

Figure 12 presents the operation with $V_{in} = 110$ V, $V_o = 100$ V, $P_o = 300$ W. This case belongs to the CCM1 operation at $d = 1$. It has been proven that CCM1 is the only mode existing when converter gain is unity. It can be seen that $\phi > \beta$, and all switches can be turned on with zero voltage regardless of load level.

The test case in Figures 13 and 14 is obtained under $V_{in} = 100$ V, $V_o = 100$ V, with the converter gain d increased to 1.1. The operation in Figure 13 is also the CCM1 operation for a full load. When the load level is reduced to near half load, the JCCM operation is observed, which is shown in Figure 14, in which the resonant current is in phase with v_p .

Both DCM and CCM3 can be observed only at high converter gain and low load. Shown in Figure 15, the waveforms of CCM3 are captured at $V_{in} = 90$ V and 30% load. The resonant current in the last interval of each half HF period is quite small, whose magnitude is design dependent. The transformer voltage during the small current duration is zero for CCM3.

When the input voltage is reduced to 80 V (*i.e.*, $d = 1.4$) and the load level is 15%, the converter is working in DCM, as shown in Figure 16. In Figure 16a, two things are found not to match the theoretical prediction depicted in Figure 7. One is that the resonant current does not remain zero for a small duration in each half period. The other is that the transformer voltage is not stable in the expected zero-current duration. The main reason is due to the non-ideal HF transformer and

other HF-sensitive parasitic components (such as the parasitic capacitance of the diodes), which has resonance with the series resonant tank during the interval, which is supposed to have no current theoretically. The new resonance current itself is small, which has less contribution to the power transfer. However, the small resonant current with higher frequency oscillation results in the fluctuation of voltage across L_s . Thus, although V_{Cs} is almost clamped as constant (shown in Figure 16b), v_T still has a large variation. To reduce such deviation, improvements on the PCB design and device specification should be helpful.

The operation waveforms of CCM2 at $d < 1$ are illustrated in Figures 17 and 18, which are measured with the maximum input voltage $V_{in} = 120$ V. For the 90% load test shown in Figure 17, phase-shift ϕ is positive and almost the same as β , so that the transformer voltage v_T is a square wave and in phase with the resonant current. In the 40% load condition of $d < 1$ shown in Figure 18, ϕ has to be reduced to negative to maintain the output voltage as constant.

5. Conclusions

In this paper, operation modes of a secondary-side phase-shifted resonant converter are identified and discussed deeply. It is pointed out that different operation modes result from the variations of converter gain, load level and the converter specifications. Equivalent circuit and subintervals in each operation mode are given and explained in detail. The steady state of all continuous current modes is analyzed uniformly using FHA. Through the evaluation of resonant current at different switching points, the soft switching conditions of all switches can be found. The boundary condition between different modes is identified, too, and illustrated with some examples graphically. The steady-state analysis results are also utilized to give a fast design of the secondary-side phase-shifted resonant converter. A lab prototype converter is then built based on the design. Different predicted operation modes are then verified through actual measurements on the prototype converter. Similar to the non-resonant version [23], this converter is attractive for a high output voltage, low current application. Due to many operation modes being available in this converter, it is not easy to make an optimized design. This work can be used as a theoretical reference for engineers to make better use of the converter or a particular application through a customized design.

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